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An Improved High-Frequency Radiotelephone System Featuring Constant Net Loss Operation

By J. M. FRASER, H. H. HAAS and M. G. SCHACHTMAN

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A system is described in which a special type of syllabic compandor and other features offer an improved solution to the problems of using HF radio facilities for overseas telephone trunks. A major disadvantage of existing systems is the use of voice-operated gates to prevent singing and echo effects. These gates interfere with the free flow of conversation. The new system allows stable full-duplex operation as in conventional telephone circuits because a nearly constant circuit loss is maintained between the trunk terminals.

Significant improvements in circuit quality result from full-duplex operation and from a reduction in the effects of radio noise and interference by compandor action. Preliminary results of a field trial between New York and Buenos Aires have been highly favorable.

I. INTRODUCTION

The first transatlantic telephone cable was laid in 1956. For almost thirty years prior to that time, transoceanic telephone communication was provided almost exclusively via high-frequency radiotelephone circuits. Numerous submarine cables have since been laid, and HF radio is no longer used for telephone service over routes of heaviest traffic, e.g., New York-London, where large numbers of cable and satellite circuits exist. Many long transoceanic routes, however, are

not served by submarine cables or satellites; direct telephone service between the United States and forty foreign countries and areas (see Fig. 1) is still provided by means of HF radio facilities exclusively.

Although a change in the composition of transoceanic telephone facilities is expected as satellites become more fully utilized, it is likely that the HF services will hold their own for economic reasons, particularly for small circuit groups.

For some time the HF radio spectrum, a rather limited resource to begin with, has been nearing the point of saturation. Thus, as new submarine cable and land* facilities make it possible to suspend existing HF routes, the frequencies are reassigned to provide service to new areas, typically to the developing nations of the world. Thus, the total number of HF circuits is expected to remain about the same well into the communication satellite era.

Since high frequencies propagate via the ionosphere, which is in a continual state of change, HF transmission is highly variable and requires special equipment and procedures to cope with the problems created by this variability. At times, HF signals (say, 4 to 27 MHz for overseas service) propagate easily over great distances; at other times, the medium will not propagate a given frequency at all. Any condition between these extremes is possible.

The ionospheric medium is a difficult and erratic one, subject to many propagation anomalies. Moreover, unique interconnection problems exist between point-to-point HF radiotelephone circuits and conventional land plant. Because of the unstable nature of the HF radio transmission medium, the combination of the 2-wire part of the land plant and a "4-wire" radio circuit presents problems that are not encountered, for example, in 4-wire cable or microwave radio circuits. Any interconnection of 4-wire and 2-wire circuits creates potential singing and echo problems. In the land plant, this problem is solved by operating the circuit at a net loss sufficient to prevent singing and to render echoes negligible.[1] Because such circuits are stable, any allowance for variations in loss is small, and the mode of operation may be characterized as "constant net loss." If HF radiotelephone circuits were operated in this mode without special controls, the allowance for variations would, in general, have to be rather large. If the design net loss were sufficient for stability with minimum transmission loss in the radio portion of the circuit, even the average loss would result in low

* HF radio is widely used for intracontinental communication in some less developed parts of the world.

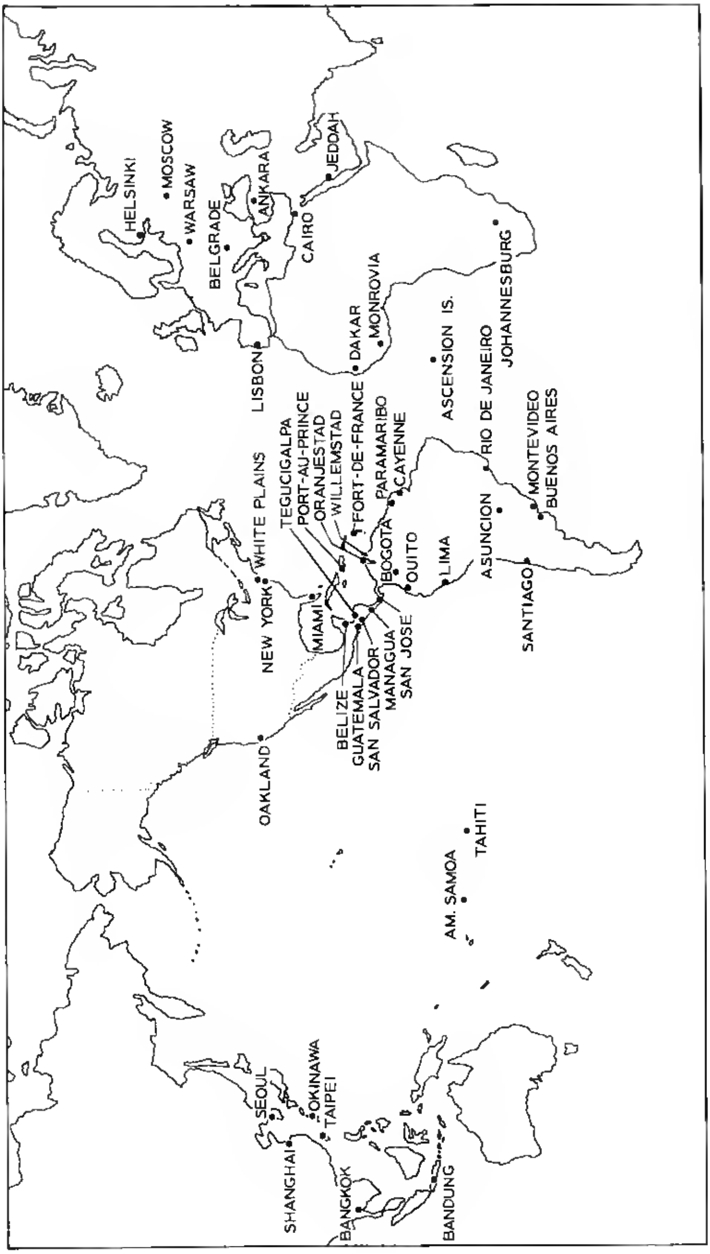

Fig. 1 — Bell System direct radiotelephone service (January 1, 1967).

received volume. The high loss occurring during deep fading would, at times, make the received volume inaudible. This would be a very inefficient mode of operation.

Input speech volumes from connecting plant vary over a wide range, at least 1000:1 in power. Linear transmission of this full range is an inefficient use of radio transmitter load capacity.

Because signal-to-noise and signal-to-interference ratios are frequently marginal, fairly high power is required for reasonable reliability. For example, at least a 4-kilowatt output might be required on a given path for adequate reception of the weakest talker. Then a 4-megawatt transmitter capacity would be required in a linear system for the strongest talker. The use of 4-megawatts perhaps one percent of the time would be uneconomical. In addition, the unnecessary radiation of excessive power would aggravate interference problems. As a counterexample, if 4 kilowatts were the maximum available to the strongest talker, the weakest talker would develop an output of only 4 watts and might be lost in the noise. These examples make it obvious that in order to minimize transmitter power and still accommodate the large range of speech volumes, some form of volume regulation is necessary. Volume regulation, which varies the gain of the circuit, compounds the problem of singing stability and echo effects discussed above.

The basic HF radiotelephone transmission problem may be sum-

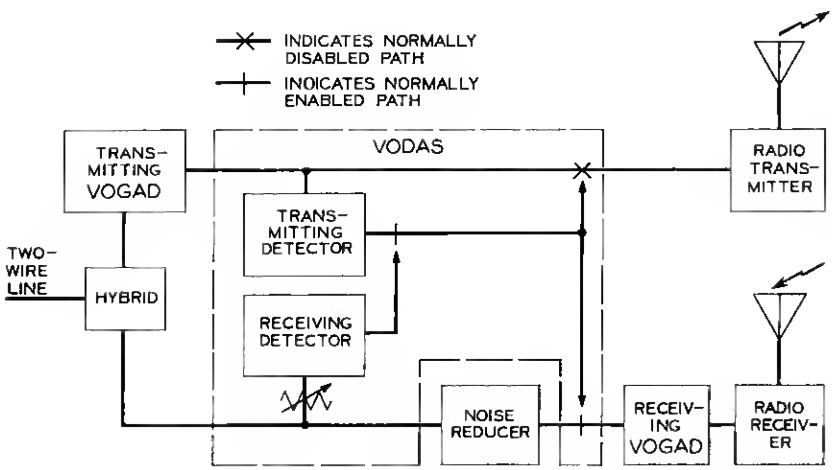

Fig. 2 — Conventional VODAS terminal.

marized at this point as one of continuously maximizing the signal-to-noise ratio under varying conditions while at the same time preventing a net gain around the circuit that would allow singing to occur. The conventional solution to this problem[2] comprises at least the following two elements:

(*i*) VODAS, or Voice Operated Device Anti-Singing, which makes the circuit one-way at a time. With this arrangement, singing and talker echoes are avoided.

(*ii*) VOGAD, or Voice Operated Gain Adjusting Device, which loads the radio transmitter efficiently.

In one form or another, these elements have been widely used throughout the world from the beginning of the service up to tbe present time.

Fig. 2 is a simplified diagram of the radio terminal equipment used in overseas gateways of the Bell System. In addition to the transmitting VOGAD and VODAS, there is a receiving VOGAD and a "noise reducer." Briefly, the terminal operates as follows. The receiving path is normally enabled and the transmitting path disabled. Outgoing speech, regulated to constant volume by the transmitting VOGAD, operates the transmitting speech detector, which enables the transmit path and disables the receiving path. When the speech train ends, the VODAS reverts to the receiving mode. The receiving VOGAD provides fading regulation of incoming speech, which has already been coarsely compensated by automatic gain control in the receiver. The noise reducer is a voice-activated expandor capable of moderate noise and interference reduction in the speech gaps if radio conditions are at least fair. The receiving detector, by inhibiting the transmitting detector, prevents echoes of the received speech from switching the circuit to the transmitting mode. The receiving detector is provided with a sensitivity adjustment that may be set by a "technical operator."

Although its wide use for a long period attests to the soundness of the present terminal, a number of disadvantages have become apparent. Foremost is the fact that even under perfect transmission conditions, a one-way-at-a-time circuit inhibits the smooth flow of conversation; in fact, when double-talking occurs, a significant amount of speech may be lost. This effect is sometimes called "lock-out". Also, because of the difficulty in differentiating between speech and noise, some "clipping" by the speech detectors occurs under the best conditions, particularly of the initial portions of speech having low energy. VODAS designs employing a receiving detector for echo protection are particularly vulnerable to operation on high received noise or in-

terference. Such false operation is called "lock-up" and makes the circuit completely unusable as long as it persists. Frequent adjustment of the receiving detector sensitivity and other controls is required if transmission quality is to be optimized with changing radio conditions.

II. AN IMPROVED RADIOTELEPHONE SYSTEM

The HF radiotelephone transmission problems can be solved by a method other than the conventional one described. Experimental systems based on general principles known for some time[3, 4] recently have been investigated by the Bell Telephone Laboratories and others.[5, 6, 7, 8] The major disadvantages of the existing system are eliminated and significant improvements are realized in the new type of system. This paper describes the experimental system investigated by the Bell Telephone Laboratories, called the constant net loss (CNL) system.

The principles of the CNL system may be explained with reference to Fig. 3, which illustrates a circuit equipped at both ends with a new type of terminal. Input speech is applied to a "complete compressor," which gives constant output volume over a wide range of inputs. The compressor smooths out the syllabic changes in speech loudness in order to fully load the radio transmitter and to optimize the received signal-to-noise ratio on a syllable-to-syllable basis. Thus, it is more effective than the slow-acting VOGAD, which regulates differences between talkers but does little to the syllabic variations of a given talker. The control signal from the first stage of the compressor, a low-frequency (0-100 Hz) analog of the syllabic variations of the input speech, feeds the two stages to give complete compression as explained in Appendix A. In addition, the control signal frequency-modulates a subcarrier, which is transmitted over the radio link along with the compressed speech, but in a separate narrow-band channel. Fig. 4 shows the frequency allocations of a single voice channel with its control channel.

The control signal is used at the receiving end to control an expandor, which restores the original variations in loudness. The FM control channel has sufficient margin to assure that even under severe noise, interference, and fading conditions, the expandor will properly "track" the compressor. The blocks labeled LOG and ANTILOG in the control channel comprise an instantaneous compandor[9] (nonlinear compressor and expandor) whose function is to make the control signal less susceptible to noise and interference in the control channel.

Although restoration of the original speech loudness variations by

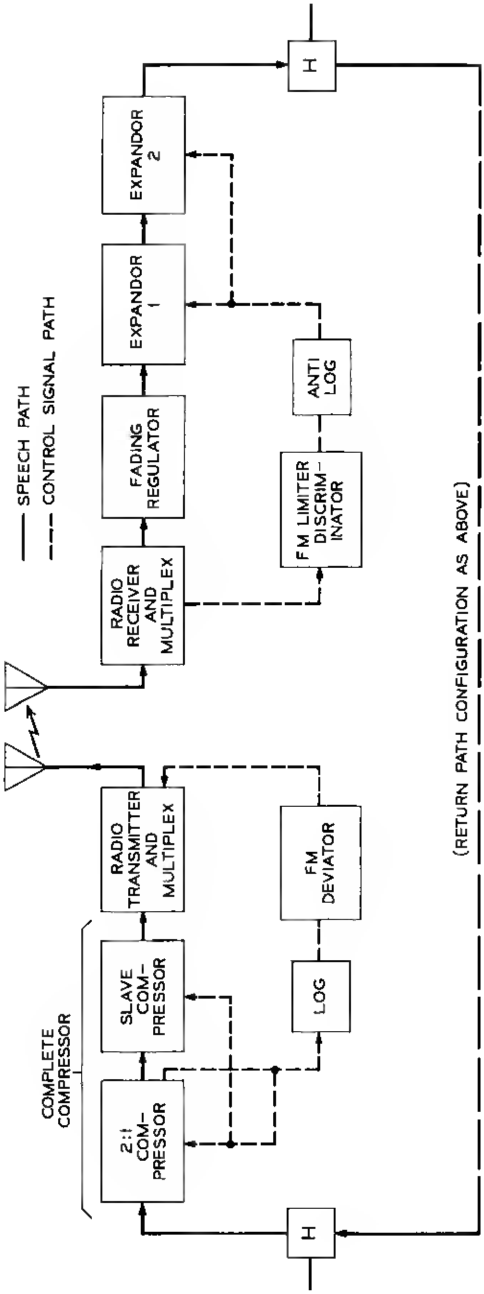

Fig. 3 — Constant net loss radiotelephone system (simplified).

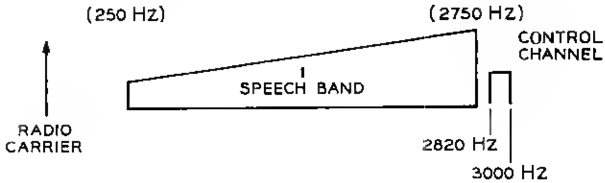

Fig. 4 — One-channel operation.

an expandor controlled from the sending end is of some benefit in itself, it has two other more essential purposes: (*i*) It maintains an overall constant net loss by making the expandor loss track the compressor gain; (*ii*) it very effectively mutes radio noise and interference during the gaps between syllables when the expandor loss is a maximum. A substantial subjective improvement accrues because the noise is reduced when it is most noticeable. Expandor tracking error caused by noise or interference in the control channel is minimized by the narrow bandwidth, FM carrier power, and other parameters of the control channel. The complementary compressor/expandor action via the control channel results in constant loss in the absence of fading. A stable loss around the outgoing and return paths allows the circuit to be set up "full-duplex" without a VODAS.

In the CNL terminal, fading variations that remain after the action of the receiver automatic gain control are absorbed by the fading regulator preceding the expandor. The fading regulator (which is also a syllabic complete compressor) operates independently of the control channel and somewhat more slowly than the transmitting compressor. The dynamic actions of the transmitting compressor, fading regulator, and expandor are such that, although gains and losses of these devices and of the transmission medium vary within the circuit, the net overall loss is approximately constant.

Since the CNL system has no VODAS with its inherent problems of clipping, lock-out, and receiving detector lock-up, a smoother flow of conversation results. Its compandor action and fading regulation are more effective in optimizing S/N (signal-to-noise) and S/I (signal-to-interference) ratios than the uncoordinated actions of the VOGADs and noise reducer of present terminals. An additional advantage is that no in-service adjustments of the terminal by technical operators are necessary, as is the case in the present terminals. A summary of the principal characteristics of experimental CNL systems is given in Table I.

TABLE I—SUMMARY OF CNL HF RADIOTELEPHONE TERMINAL CHARACTERISTICS

System Capacity	Four speech channels plus four control channels can be accommodated on one radio carrier in conjunction with conventional 12-kHz radio and channelizing equipment.
Transmitting Compressor Operating Range	The range of input power over which compressor output is essentially constant is +8 dBm0 to −32 dBm0. Below −32 dBm0, the compressor acts as a linear amplifier; above +8 dBm0, some clipping of a sinewave occurs.
Speech Bandwidth	The nominal speech bandwidth is 250 to 2750 Hz for channels next to the carrier, 250 to 2950 Hz for outboard channels.
Control Signal Allocations and Power	(*i*) The FM control signal associated with an inboard channel occupies an 180-Hz band centered at 2910 Hz. With no system input, the subcarrier frequency is 2850 Hz; maximum test tone input (+8 dBm0) deviates the subcarrier upward 120 Hz to 2970 Hz. (*ii*) The outboard channel, 180 Hz wide, is centered on 3140 Hz, and the corresponding subcarrier frequency excursion is from 3200 Hz downward to 3080 Hz. (*iii*) The FM subcarrier power in the composite signal is applied to the radio transmitter at a level 16 dB below its peak envelope power rating and 6 dB below a test tone in the speech channel, the test tone being within the constant output range of the transmitting compressor.
Time Constants	The transmitting compressor has an attack time of 3 msec and a recovery time of 13.5 msec. The overall effective RC time constant of the compandor system is 20 msec.
Fading Regulator	The fading regulator maintains essentially constant output over an input range 20 dB below and 10 dB above the nominal no-fade level. The attack and recovery times are 12 msec and 54 msec, respectively. These values are 4 times the corresponding transmitting compressor time constants.
FM Control Signal Deviation/Loss Ratio	The deviation/loss ratio is the transfer constant that relates the deviation of the FM subcarrier to the expandor loss variations and is the best measure of the susceptability of the control channel to noise and frequency instability. The constant is 2 Hz per dB.

The CNL terminal is compatible with existing channelizing equipment, privacy devices, radio transmitters, and receivers. The system uses no more bandwidth overall than present systems. Although the FM channel shares bandwith with the speech, the loss of speech bandwidth is slight. This is possible because of more efficient use of the available bandwidth by means of sharper filter cut-offs. In conjunction

with conventional 12-kHz radio and channelizing equipment, a full complement of four speech channels plus four control channels can be accommodated (Fig. 5).

The control signal must share the available power of an existing transmitter with the compressed speech. Satisfactory performance is expected with control subcarrier magnitudes (see Table I) such that the total transmitter load is not significantly increased.

There are several reasons for using FM to transmit the control information. The foremost reason is that limiter action provides level compensation, making the control signal insensitive to fading (as regards purely level, rather than S/N variations). Also, bandwidth can be traded for S/N advantage. In addition, with FM, the dc com-

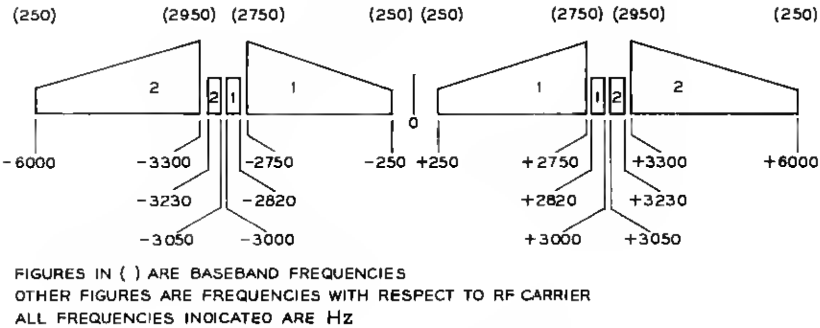

Fig. 5 — Four-channel operation.

ponent of the control signal may be preserved.* Finally, the narrow-band control signal is similar in many ways to narrow-band telegraph signals which have been transmitted successfully via HF radio for many years using FM or related methods. It is too early to predict the extent to which diversity would be useful or necessary in a CNL-type radiotelephone system, although it is widely used with HF radio telegraph systems.

A successful trial of experimental CNL terminals has been conducted between New York and Buenos Aires by the American Telephone and Telegraph Company with the cooperation of the foreign correspondent. Conventional terminals on one circuit of a regular four-channel system were replaced with CNL terminals. A large number of test and demonstration calls were made including some comparisons

* The control signal is derived by rectifying the speech output of the first compressor stage; it is therefore unipolar and has a dc component.

between the CNL circuit and a conventional circuit operating in the same radio system. Radio conditions on this path generally are fair to good, neglecting propagation outages. The participants in these tests were largely persons familiar with both cable and conventional radio overseas circuit performance. In the judgment of most of the participants, the CNL circuit quality approached that of a submarine cable circuit because of the compandor action and lack of VODAS impairments.

Following the above demonstrations in July, 1966, the CNL-equipped circuit began an extended period of commercial service. Traffic records indicate that the single CNL circuit carried one quarter of all calls on the New York-Buenos Aires route, which has a total of 12 circuits. The CNL circuit handled more calls than the next two most active conventional circuits combined. In the opinion of operating personnel, calls that would have encountered considerable impairments or operating difficulties with conventional terminals were handled without customer complaint. During periods of poor radio conditions, the CNL circuit was frequently "commercial" when some, if not all, of the conventional circuits were unusable.

Similar experimental terminals have been developed and tested by the British General Post Office (GPO) and French Postes Et Telecommunications (PTT). The British and American Administrations submitted reports of their work on the new type of terminal to the plenary meeting of the International Radio Consultative Committee (CCIR) in Oslo in the summer of 1966. This report included essential design parameters and operating conditions, and was submitted as an initial contribution toward adoption of international compatibility standards for the new system.

III. DESCRIPTION AND ANALYSIS OF IMPROVED SYSTEM

3.1 *System Application—Overall Block Diagram*

The block diagram Fig. 6 illustrates an arrangement of the experimental CNL equipment that was installed on a working radio circuit between Buenos Aires and New York. The equipment shown as shaded boxes is the same as is used with the present terminals. Other arrangements of signaling, privacy, channel shifters, and filters are possible; the best arrangements will depend on the characteristics of the specific equipment to be used on a given circuit.

One independent sideband (ISB) HF transmitter normally carries

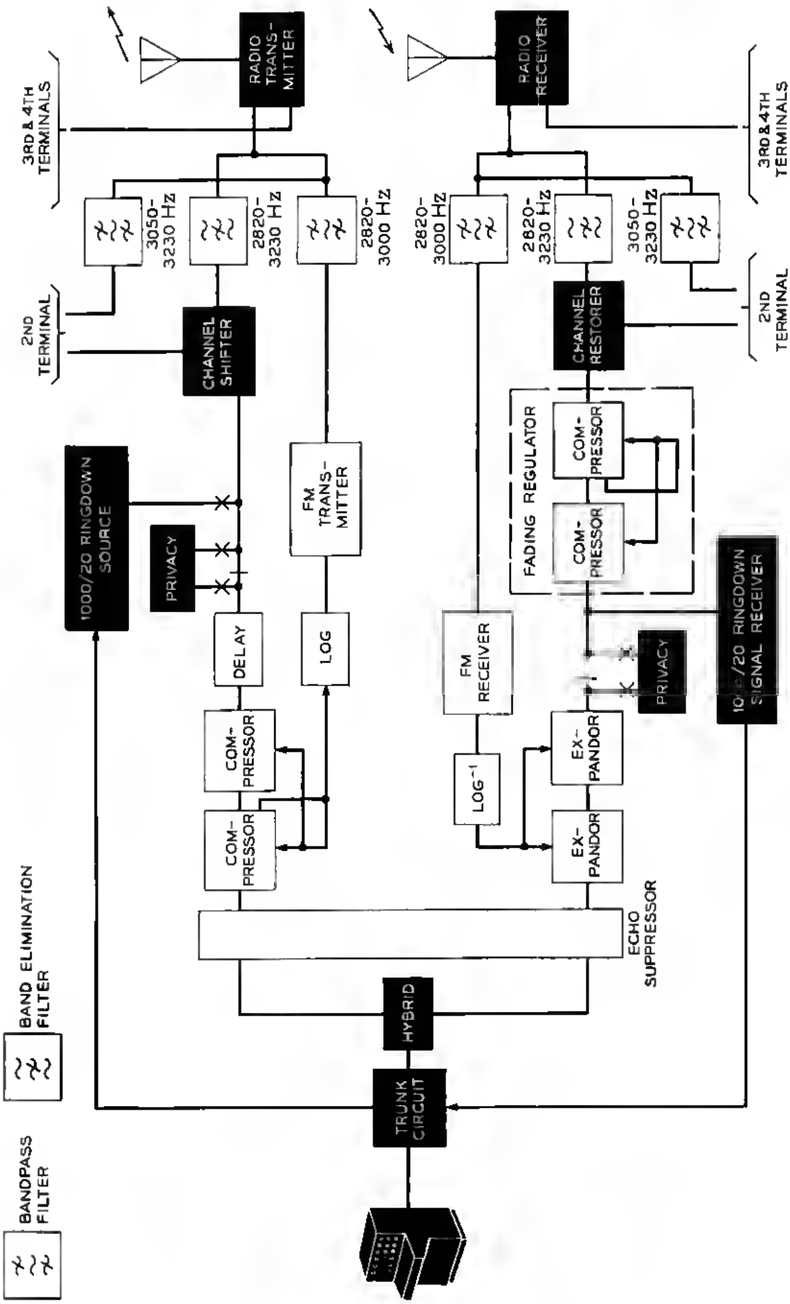

Fig. 6 — System block diagram.

a maximum of four channels, two on each sideband. The particular arrangement of Fig. 6 was designed, among other reasons, to facilitate building up multichannel CNL systems. In such systems, the voice outputs of two separate terminals are combined in a channel shifter; a single band-elimination filter at the output of the shifter produces a slot to accommodate two control channels. The control subcarrier of each terminal is transmitted through a bandpass filter of the appropriate frequency; the subcarriers from two terminals are then combined with the output of the band-elimination filter. The combination of the two speech and two control channels is applied to one sideband of an ISB transmitter. The ISB transmitter has provision for two additional circuits on the other sideband, as shown in Fig. 5. Since only one circuit was equipped with CNL terminals in the trial installation, the second, third, and fourth terminals indicated on Fig. 6 were conventional VODAS terminals.

The filter and channel restorer arrangement of the receiving side is essentially a duplicate of the transmitting side, with a single band-elimination filter for one pair of speech channels and bandpass filters for each of two control channels. The attenuation required to control crosstalk between the speech and the control channels is provided by the combination of transmitting and receiving filters. It will be noted from Fig. 6 that the control channel was combined with the speech channel after the channel shifter and split before the channel restorer. This approach was used so that the characteristics of particular channel shifters and restorers would not affect the control channels.

Signaling for the trial installation was based on manual operation of the radio telephone circuit, with a ring-down signal of 1000 Hz modulated at 20 Hz inserted on the radio side of the terminal after the privacy. A 1000/20 signaling receiver, bridged after the fading regulator and before the privacy and expandor on the receive side, operated the appropriate switchboard equipment. Suitable trunk, monitoring, and control circuits similar to those found in VODAS terminals were provided.

The trial installation includes a flat delay unit to equalize the delay of the speech path with respect to the separate narrowband control path. (A discussion of delay equalization is given in Section 3.2.4.)

Terminals of the CNL type are inherently full-duplex and all equipment exclusive of the 4-wire terminating set, or hybrid, is arranged on a 4-wire basis. Fig. 6 shows connection to a 2-wire switchboard; 4-wire switchboards are also used. Privacy equipment must be 4-wire to preserve the two-way nature of the CNL principle. Either the inversion

or split-band type of privacy equipment may be used. In the split-band privacy, the speech is separated into five 550-Hz bands and transposed in frequency in any arrangement that provides reasonable privacy. With the CNL-type terminal, it may be necessary to restrict the possible transpositions to insure that the control channel slot is always taken out of the speech around 3000 Hz in order to minimize the effect of removing this slot from the speech band.

An echo suppressor is required on all long circuits because of the inherent propagation time. The receiving split-type echo suppressor used on the Buenos Aires-New York circuit has a speech gate only in the transmit leg; the gate blocks transmitted speech only when speech of greater magnitude is present in the receiving leg. This type of suppressor was preferred because of its break-in properties and also because it tends to reduce reradiation of echoes.

3.2 *Analysis of Factors Affecting Performance*

3.2.1 *General Considerations*

As mentioned previously, allocations of available bandwidth and transmitter power must be made in a CNL-type system, in which the compressed speech and the information to control the expandor are transmitted via separate channels. The objective in making this allocation is to obtain adequate margin against noise and interference in the control channel without encroaching substantially on the speech bandwidth and power. While the speech bandwidth is not critical, it is not a simple matter to determine what bandwidth should be allocated to the control signal. The procedure used in selecting the parameters of the experimental CNL system was to make a somewhat arbitrary allocation of bandwidth and power, then to investigate what further trade-offs could be made within these constraints.

The control channel bandwidth selected was 180 Hz. The net speech bandwidths obtained were as indicated on Fig. 5 and Table I. The impairment due to loss of about 200 Hz of high-frequency speech energy of the inboard channel is slight. There is no loss of speech bandwidth to the outboard channel because the control channel in this case lies in what was originally guard space (between 3 and 3.25 kHz).

In the experimental CNL system, the available power per channel was divided in the ratio 1:4, i.e., the FM subcarrier power was 6 dB below a 0-dBm0 voice-frequency test tone. The total load, including speech and subcarrier, was then only about 1 dB greater than the

speech power alone. It would be undesirable to put so much power in the subcarrier as to necessitate a reduction in the speech drive to the transmitter and a consequent reduction in signal-to-noise ratio at the receiver.

For further discussion of factors affecting performance of the improved system, attention is drawn to the fact that the transmitting compressor and receiving expandor together comprise a syllabic compandor, which is closely related to more conventional devices of this class. In these devices, as well as in the CNL system, the transmission gain (or loss) is varied at syllabic rates in order to gain advantage against noise and interference. The conventional syllabic compandor has a compression ratio of 2:1, and uses the residual speech envelope of the compressed signal to derive the information with which to control the expandor. Thus, it does not require a separate control channel, although there are circumstances where such an arrangement could be advantageous. For example, a control channel could improve tracking of such a compandor in the presence of high noise at the expandor input.

Speech can be transmitted through a conventional compandor with low distortion using no more bandwidth than that of the original speech as long as the rate of gain variation, or "speed," does not exceed syllabic rates. Hence, the bandwidth, within limits, is not a consideration in establishing the speed of a conventional compandor.

The following general requirements govern the speed of a syllabic compandor. (*i*) *Compressor speed:* a fast compressor is more effective than a slow compressor in raising weak speech syllables with respect to the noise before transmission. If this action is too abrupt, however, significant distortion is created. (*ii*) *Expandor speed:* the speed of the expandor must match that of the compressor. If it is too slow, the expandor may mutilate the initial parts of a syllable, or its loss may not be fully inserted during the gaps so that the noise reduction effect of the expandor suffers.

Within the range of syllabic rates, the speed of a conventional compandor may be varied over a fairly broad range and still satisfy the foregoing requirements. Furthermore, the compandor noise improvement, i.e., how well the compressor picks up weak speech before transmission and the expandor mutes the noise between syllables, is not highly sensitive to speed.

The use of a separate control channel to control the expandor in the CNL system introduces an additional factor governing speed of the

CNL compandor; otherwise, the criteria governing speed would be the same as a conventional compandor. The additional factor is the noise performance of the control channel, which is strongly dependent on compandor speed. The speed of the compandor in the CNL system may be regarded as an independent parameter upon which two main categories of system performance depend. These are: (*i*) the control channel noise performance and (*ii*) the compandor noise improvement. The first degrades with increasing speed and the second improves with increasing speed; therefore, it is necessary to select parameters such that a reasonable balance is achieved.

The principal elements of the CNL system affecting this balance can be represented by the model of Fig. 7. The control signal is derived from the output of the first compressor variolosser by rectifying a portion of the speech voltage at this point. The unfiltered output undergoes an initial smoothing, or prefiltering, in the network R_0C_0. It then branches into two paths, one feeding back through the low pass amplifier (R_1C_1) to control both variolossers of the compressor and the other feeding forward via the control channel and the low pass amplifier (R_2C_2) to control the variolossers of the expandor. R_0C_0 band limits the spectrum of the modulating signal at the input to the control channel. The primary role of R_2C_2 is that of a post-detection noise filter. In addition to their filtering functions, the networks R_0C_0 and R_2C_2 together establish the time response of the expandor. The FM channel itself can be made to have a negligible effect on the expandor response. A unique feature of this particular configuration is that, regardless of the overall speed of the compandor, the time response of the feedback path can be matched to the feed-forward path by simply making $R_1C_1 = R_2C_2$; dynamic tracking of the compressor and expandor is thus assured.*

R_1C_1 and R_2C_2 are single-pole feedback networks of identical operational amplifiers. The bandwidth of these elements, their time responses, and the equivalence between bandwidth and time response may be analyzed in straightforward manner. The 3-dB bandwidth frequency of an RC low-pass characteristic is

$$f = \frac{1}{2\pi\ \mathrm{RC}},$$

* The representative spectra shown in Fig. 7 are illustrative of the significant bandwidths present at several points in the CNL terminal. Note that the noise bandwidth at the expandor variolosser is f_2 while the effective bandwidth of the control signal paths is less because of the influence of R_0C_0 in tandem with R_1C_1 or R_2C_2.

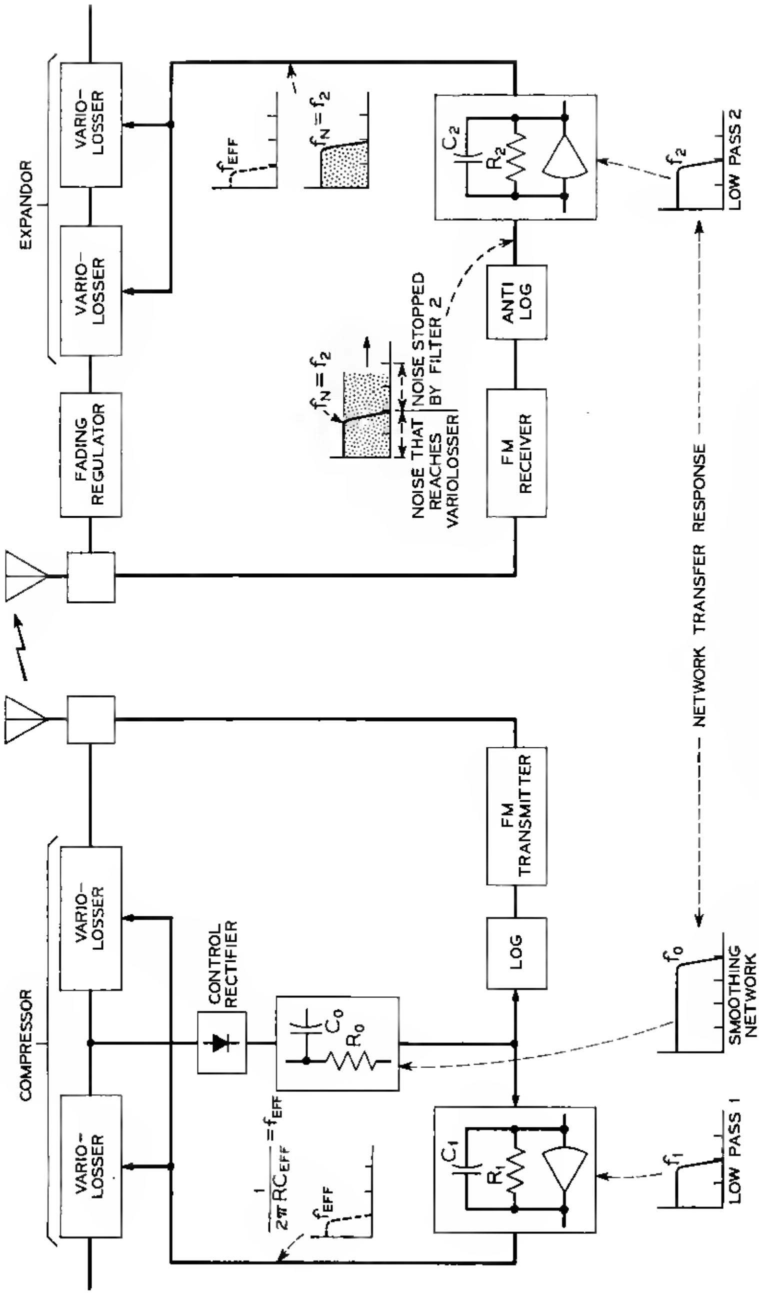

Fig. 7 — Time constant determining elements of CNL system.

from which a direct connection between speed of the compandor and noise performance of the control channel will now be apparent. That is, as the speed of the compandor is increased (by decreasing R_1C_1 and R_2C_2), f_2 widens to admit more noise to the control lead of the expandor. The fluctuation of expandor loss about its nominal value produced by random noise is, in effect, low-frequency amplitude modulation of the speech. This imparts a unique fluttering or "gritty" quality to the speech, which is a source of impairment over and above the effect of additive noise falling in the speech band. Interfering signals falling in the control channel also produce loss deviations, but of a less random nature. A limitation on the noise reaching the expandor via the control channel imposes a corresponding limitation on $f_2 = 1/(2\pi R_2C_2)$. R_2C_2 strongly influences the resulting speed of the compandor, and therefore the compandor noise improvement. Both the effect of noise in the control channel and the compandor noise improvement are for the most part ultimately subjective. The task of finding an objective basis for evaluating these effects and analyzing them in detail is taken up in the following sections.

In Section 3.2.2 and related Appendix B, control channel noise performance is analyzed, a criterion of performance is established, and subjective limits are given.

In Section 3.2.3 the speed of the CNL compandor is discussed on a quantitative basis; relationships between time response and bandwidths are given. The related problem of dynamic tracking and the choice of operating ranges are dealt with in Section 3.2.4.

In Sections 3.2.5 and 3.2.6, respectively, the expandor noise improvement and the efficacy of the complete compressor in loading the transmitter are discussed quantitatively, both as a function of compandor speed.

3.2.2 *Noise Performance of the Control Channel*

The performance of the control channel has two aspects: (*i*) "normal" operation when the subcarrier is above a certain noise threshold and (*ii*) the breaking threshold below which the channel breaks down.

The breaking threshold is reached in the FM channel of the CNL system when noise or interference at the discriminator unit is so strong relative to the subcarrier that the polarity of the signal is too often reversed. Under these conditions, impulsive noise currents are generated at the discriminator output; they drive the expandor to its extreme loss or gain. As the threshold is approached, the impulsive noise

currents punch "holes" in the speech or produce annoying intervals of excessive loudness.

The threshold occurs with random noise when the subcarrier-to-noise ratio is in the vicinity of 10 dB on an rms basis. This ratio is a function of the line, or predetection, bandwidth and the subcarrier power. In the case of the experimental CNL system, the line bandwidth is 180 Hz, and the subcarrier power is 6 dB below the mean compressed speech power. Therefore, with the assumption that the noise is flat across the band, breaking occurs when the ratio of compressed speech to noise in the 2500-Hz speech band is about 5 dB.

The bandwidth occupied by the FM sidebands must not exceed the 180-Hz bandwidth allocated. It is a complex function of the amplitude and spectral distributions of the modulating signal and the frequency swing of the FM subcarrier. To further complicate matters, the modulating voltage is not symmetrically distributed and its spectrum is modified by the nonlinear LOG circuit preceding the frequency modulator. It was experimentally determined that the FM spectrum resulting from a control signal rolloff at 100 Hz (3 dB down) and a peak-to-peak swing of 120 Hz was in accord with the 180-Hz line bandwidth. The 100-Hz roll-off occurs in the control rectifier filter, represented symbolically by R_0C_0 on Fig. 7. The 120-Hz swing divided by the expandor loss range (60 dB, as discussed in Section 3.2.4) establishes the deviation/loss ratio. It relates frequency errors of the FM subcarrier to the system loss error. These errors are due to noise and other mechanisms. The ratio enters into the calculations of Appendix B.

The magnitude of system loss fluctuations above the breaking threshold due to noise in the control channel has been calculated by means of an expression derived in Appendix B. The derivation assumes that the noise is random and flat across the band common to both the compressed speech and the subcarrier. Although a more usual situation in HF reception is a mixture of noise and tone-like interfering signals having a variety of characteristics together with selective fading, the simple model nevertheless gives useful results.

Equation (40) from Appendix B is

$$\begin{array}{c}\text{loss fluctuation magnitude}\\ \text{(dB peak-to-peak)}\end{array} = \frac{3(f_2)^{\frac{1}{2}}}{25\ \mathrm{S/N}}\left[\frac{90}{f_2} - \tan^{-1}\left(\frac{90}{f_2}\right)\right]^{\frac{1}{2}}.$$

Fig. 8 is a plot of the loss fluctuation as a function of S/N with f_2 as a parameter. S/N is the ratio (numeric) of the speech voltage to the

noise voltage in the 2500-Hz speech band and is an indirect measure of the noise falling in the control channel. Expressing results in this manner makes it possible to correlate quality judgments based on the speech-to-noise ratio of the circuit with the degree of impairment caused by fluctuating system loss. By subjective tests, the approximate magnitudes were determined at which the loss fluctuations were judged (*i*) to have just noticeable impairment and (*ii*) to be so severe as to render the circuit uncommercial. These magnitudes are indicated on Fig. 8 as cross-hatched bands.

The abcissa of Fig. 8 refers to the S/N at the input to the receiving terminal. In a circuit equipped with CNL-type terminals, a S/N of about 10 dB would be judged uncommercial and 15 dB would be judged poor but usable. By the standards normally applied to HF radiotele-

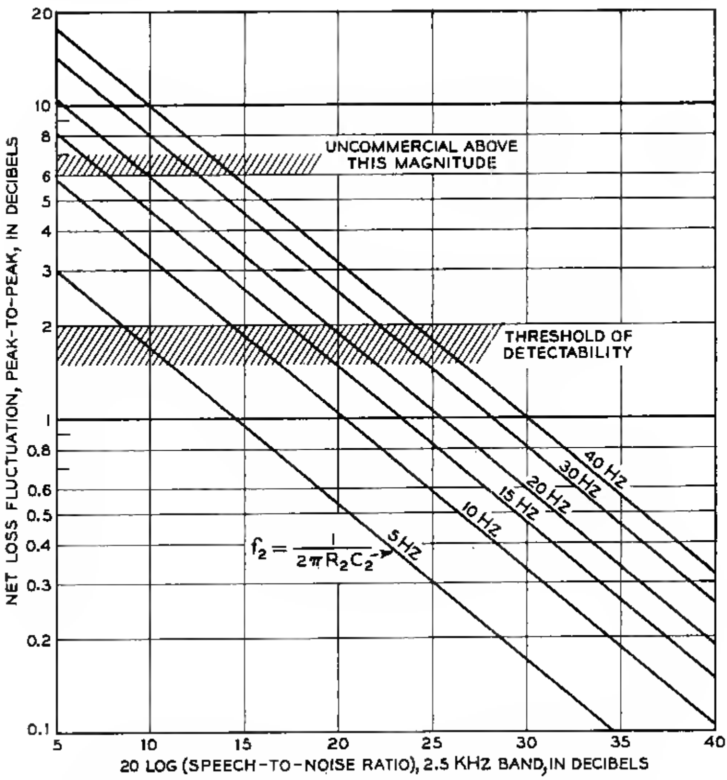

Fig. 8 — Net loss fluctuation as a function of speech-to-noise ratio with control channel bandwidth as a parameter.

phone circuits, 25 dB would be considered good. Fig. 8 shows that if f_2 becomes large, the effect of noise in the control channel would be detectable even under otherwise good conditions. On the other hand, an f_2 of 5 Hz or less provides more margin than is necessary. With an f_2 of 20 Hz, when the S/N drops to 10 dB, the loss fluctuations reach an uncommercial magnitude. This same S/N would be judged uncommercial with a noise-free control channel. Thus, with an f_2 of 20 Hz, the limit of commercial quality would be reached more or less simultaneously due to noise mixed with the speech and noise falling in the control channel. The value of f_2 adopted for the experimental CNL system was 15 Hz. It tends to be conservative in respect to loss fluctuation at the expense of compandor speed. The other speed-dependent performance factors discussed in Sections 3.2.5 and 3.2.6 change slowly with speed.

The CNL system employs nonlinear signal processing, or "instantaneous companding,"[4,9] within the control channel. This is the function of the LOG and ANTILOG circuits shown in Fig. 7. Such processing reduces the effect of noise on weak control signals. With weak talkers or during idle circuit conditions the control channel can tolerate more noise for a given requirement on expandor loss fluctuations if instantaneous companding is used. While instantaneous companding increases the bandwidth of the analog control signal, a net improvement in control channel performance results in the CNL application. The response of the LOG circuit itself is plotted on normalized linear scales on Fig. 9. The ANTILOG characteristic must be sufficiently complementary to the LOG characteristic to insure overall linearity of the control channel.

3.2.3 *Time Constants*

Before considering CNL compandor speeds, it is necessary to have a precise definition of time constants. Also, an accepted definition will permit comparison of the CNL with other types of compandors, in particular, results obtained by other organizations (British GPO, etc.) currently working on CNL-type systems. The CCITT* has developed a definition for message compandor time constants that has found wide acceptance. This definition has the virtue of being readily implemented with simple measuring techniques and is suitable for specifying time constants in the CNL compandors described herein.

The definition of attack and recovery time constants of a compandor

* International Telegraph and Telephone Consultative Committee.

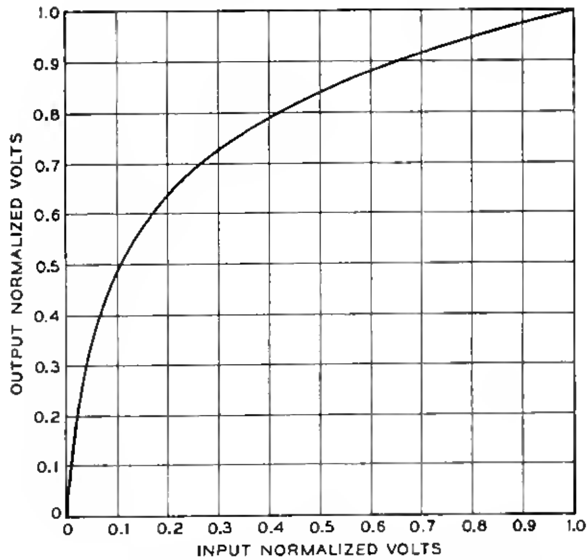

Fig. 9 — Log circuit input-output characteristic.

involves a measurement of the transient response of a 2:1 compressor. In the CNL system, the first stage is identical to the conventional compressor and the second variolosser is slaved to the first. This enables the CNL compressor time constants to be measured using the first compressor in accordance with the methods specified by the CCITT. The attack time is measured by increasing the level of the input tone by 12 dB from −16 dBm0 to −4 dBm0 and measuring the time interval between the occurrence of the step and the time when the output envelope reaches 1.5 times the final steady-state value. The recovery time is the time interval between the occurrence of a downward step of 12 dB (from −4 dBm0 to −16 dBm0) and the time when the output waveform reaches 0.75 times the final steady-state value.

For message compandors used on wire and radio relay circuits of moderate length, the recommended CCITT values are:[10,11]

$$\text{Attack time} = 3 \pm 2 \text{ msec}$$

$$\text{Recovery time} = 13.5 \pm 6.5 \text{ msec.}$$

In order to relate measured attack and recovery times of a compandor to circuit elements as an aid to design, R. O. Carter[12] analyzed a simple model of a compressor (Fig. 10) and developed formulas

for calculating attack and recovery times. In his model, the control rectifier consisted of a peak detector (diode) and a single RC smoothing network. He showed that the CCITT attack and recovery times were given in terms of the product RC (which has the dimensions of time) as follows:

$$\text{Attack time} = 0.15\ \text{RC} \quad (1)$$

$$\text{Recovery time} = 0.675\ \text{RC}. \quad (2)$$

The value of RC, since it is common to both the attack and recovery time constants, provides a single measure of the speed of the compres-

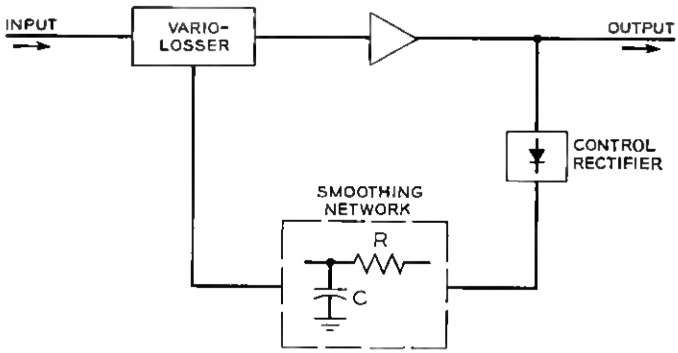

Fig. 10 — Simple model of 2:1 compressor.

sor. The bandwidth of the control signal into the variolosser is also a function of RC:

$$f = 1/(2\pi\ \text{RC}).\ * \quad (3)$$

Using relations (1) and (2), the CCITT recommended time constants of 3-msec attack and 13.5-msec recovery correspond to a speed of RC = 20 msec. The objective of the following analysis is to examine the CNL system model (Fig. 7) and find an $\text{RC}_{\text{effective}}$ for the somewhat more complicated CNL compandor.

The transient response for the compressor configuration with R_0C_0 and R_1C_1 networks was calculated approximately by divorcing the

* This relationship permits a ready transformation between factors best considered on a time (or speed) basis and factors best considered on a frequency (or bandwidth) basis. It is exact only when a simple RC network is used in the model. The bandwidth, f, corresponds to the frequency at which the control path gain is 3 dB down from the value at dc.

time-constant determining components from the nonlinear feedback circuit and solving their response to a step input. A similar step response was determined for the simple RC network of Carter's model for the time $t = 0.150$ RC, the CCITT attack time. The two responses were made equal and the attack time for the CNL compressor, t_A, was obtained by solution of

$$1 - \exp\left(-\frac{0.150\ \mathrm{RC}}{\mathrm{RC}}\right) = 0.139$$

$$= 1 + \frac{\mathrm{R_0C_0} \exp\left(-\frac{1}{\mathrm{R_0C_0}} t_A\right) - \mathrm{R_1C_1} \exp\left(-\frac{1}{\mathrm{R_1C_1}} t_A\right)}{\mathrm{R_1C_1} - \mathrm{R_0C_0}}. \tag{4}$$

The use of the step response of the networks alone to obtain an engineering estimate of the attack time was justified because the input to the speed-determining element under CCITT test conditions is essentially a step function in the time scales considered here. The speed of the CNL compressor (given by an $\mathrm{RC_{effective}}$) is determined from this calculated attack time through use of (1). Similarly, the recovery time was obtained by solution of the following equation for t_R:

$$\exp\left(-\frac{0.675\ \mathrm{RC}}{\mathrm{RC}}\right) = 0.509$$

$$= \frac{\mathrm{R_1C_1} \exp\left(-\frac{1}{\mathrm{R_1C_1}} t_R\right) - \mathrm{R_0C_0} \exp\left(-\frac{1}{\mathrm{R_0C_0}} t_R\right)}{\mathrm{R_1C_1} - \mathrm{R_0C_0}}. \tag{5}$$

This calculated recovery time can also be used to obtain $\mathrm{RC_{effective}}$ using (2). The validity of expressions (4) and (5) was confirmed by direct measurement of attack and recovery times. Expressions (4) and (5) reduce to the form of (1) and (2) when either RC is much greater than the other. When the two time constants are equal,

$$t_A(\text{for } \mathrm{R_1C_1} = \mathrm{R_0C_0}) = 0.652\ \mathrm{R_0C_0} \tag{6}$$

$$t_R(\text{for } \mathrm{R_1C_1} = \mathrm{R_0C_0}) = 1.649\ \mathrm{R_0C_0}\ , \tag{7}$$

an attack time increase of over 4 times and a recovery time increase of almost $2\frac{1}{2}$ times over the values of a single $\mathrm{R_0C_0}$, rather than a twofold increase as might be surmised at first glance. Plots of both the attack and recovery equations (4) and (5) for values of $\mathrm{R_0C_0}$ and $\mathrm{R_1C_1}$ in the range of interest are given in Figs. 11(a) and 11(b). These curves illustrate how the values of $\mathrm{R_0C_0}$ can be adjusted to yield a desired attack or recovery time when $\mathrm{R_1C_1}$ is specified.

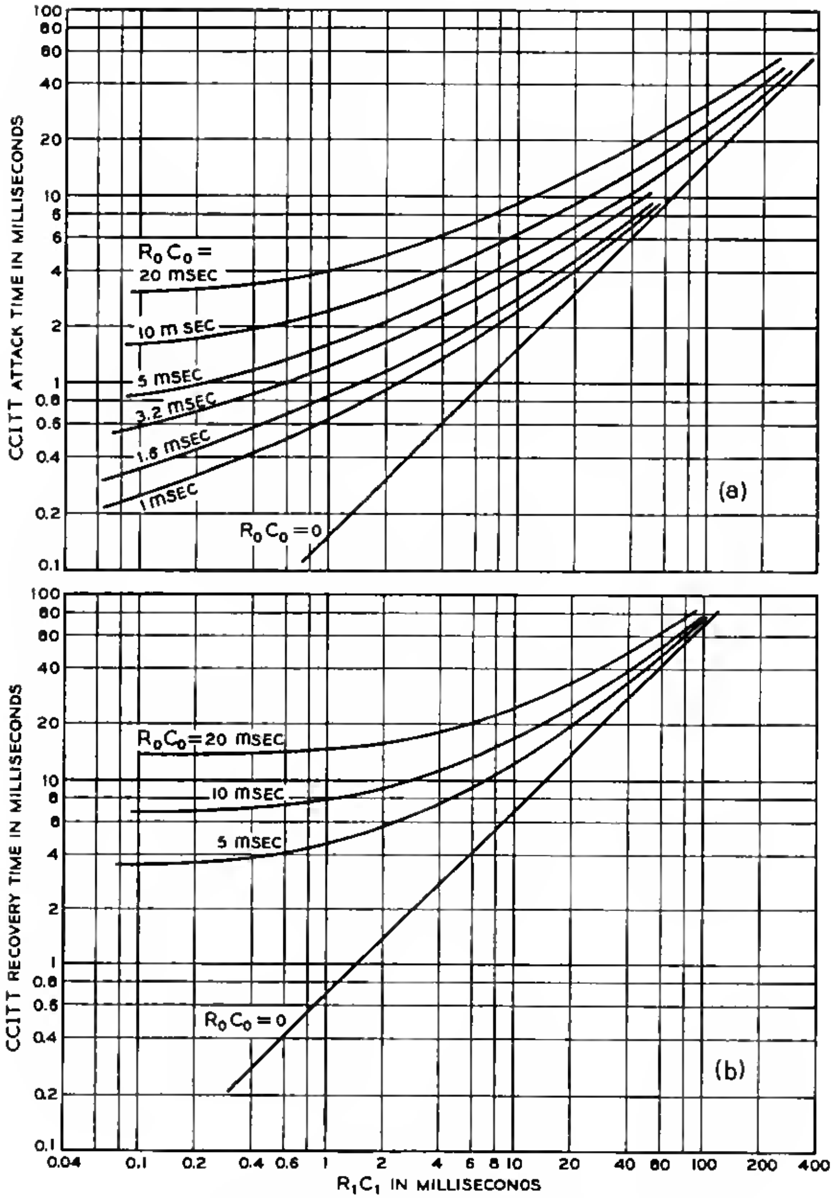

Fig. 11 — (a) Calculated attack time; (b) calculated recovery time.

In Section 3.2.2, the frequency determined by R_2C_2 was given as 15 Hz; substituting this value in (3) gives an $R_1C_1 = R_2C_2$ of 10.6 msec. The objective in selecting R_0C_0 was to yield compressor attack and recovery times recommended by the CCITT. This resulted in a choice of

$$R_0C_0(\text{attack}) = 1.6 \text{ msec}$$

$$R_0C_0(\text{recovery}) = 5.0 \text{ msec.*}$$

From Figs. 11(a) and 11(b) these values give overall attack and recovery times as follows:

$$\text{Attack time} = 3 \text{ msec}$$

$$\text{Recovery time} = 13.5 \text{ msec.}$$

The corresponding $RC_{\text{effective}}$ given by (1) and (2) was 20 msec.

Fig. 12 shows the relationship between overall speed ($RC_{\text{effective}}$) and the characteristics of the lowpass networks (R_1C_1 and R_2C_2).

3.2.4 *Dynamic Tracking and Operating Ranges*

As mentioned in Section 3.2.1, the control paths from control rectifier to variolossers pass through essentially identical smoothing networks. Therefore, the input control signals to the transmitting compressor variolossers and receiving expandor variolossers have the same transient characteristic. Matching the control signals into the variolossers insures that the expandor loss matches the compressor gain on a dynamic basis. However, an imperfect reconstruction of the speech waveform at the system output can occur if:

(*i*) the transmission time delay of the control signal differs from that of the corresponding speech signal, causing a mismatch between the instantaneous expandor loss and the envelope of the speech signal, and/or

(*ii*) the fade regulator is too fast, thus interpreting the residual amplitude variations in the output of the transmitting complete-compressor as fades and introducing further compression (without corresponding expansion).

These effects result in speech that sounds distorted and possible circuit instability if round-trip propagation delay is short.

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* The two values of R_0C_0 require that the simple model of Fig. 7 be modified to provide a control rectifier with different charge and discharge times.

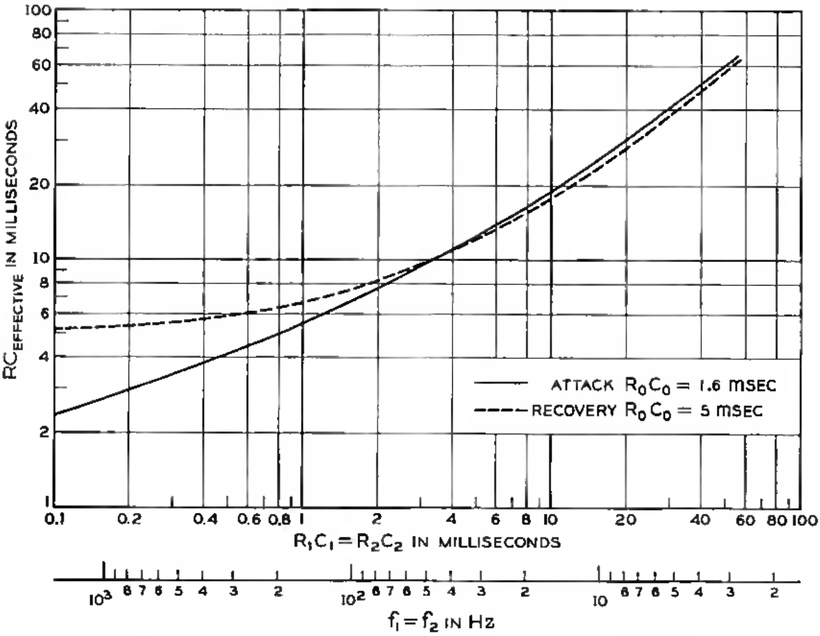

Fig. 12 — Overall speed versus low pass network time constant.

It has been found that a fade regulator RC time constant approximately 4 times $RC_{effective}$ reduces the additional compression distortion to acceptable levels without affecting performance during fading.

Flat voice-frequency delay equalization in the speech or pilot channel is used to correct the transmission time differences due to filter and privacy flat delays. For speech service, it is desirable to reduce the time difference to no more than 5 to 10 msec. For signaling and similar pulse transmission where pulses are amplitude modulated on a signaling tone, the distortion due to time differences is more severe; a 1- or 2-msec difference produces noticeable changes in pulse shape. Equalization requirements for actual signaling systems of this type would depend upon system sensitivity to pulse shape.

The need for time delay equalization is reduced if the CNL compandor is slow. If $R_1C_1 = R_2C_2$ is large compared with the time difference, the rate of change of control current into the variolossers is slow in any interval corresponding to the transmission time difference. The expandor loss mistracking errors will therefore be small since the slope of the loss curve as a function of time is low.

The operating ranges of the variable gain elements of the CNL system are interdependent since the sum of the variable gains and losses, including medium variations, results in an approximately constant value under both dynamic and static conditions. This is essential for system stability and tracking. The transmitting compressor range is 40 dB; the fading regulator downward range is 20 dB and the upward range is 10 dB; the expandor range is 60 dB, and is the sum of the transmitting compressor range and the fading regulator downward range (see Fig. 13). The operating ranges chosen were a reasonable compromise based upon expected speech volume and fading ranges, as well as practical considerations.

3.2.5 *Syllabic Compandor Quieting Versus Speed*

The speed of the compressor affects the degree of compandor quieting of system noise heard by the subscriber. A limited subjective com-

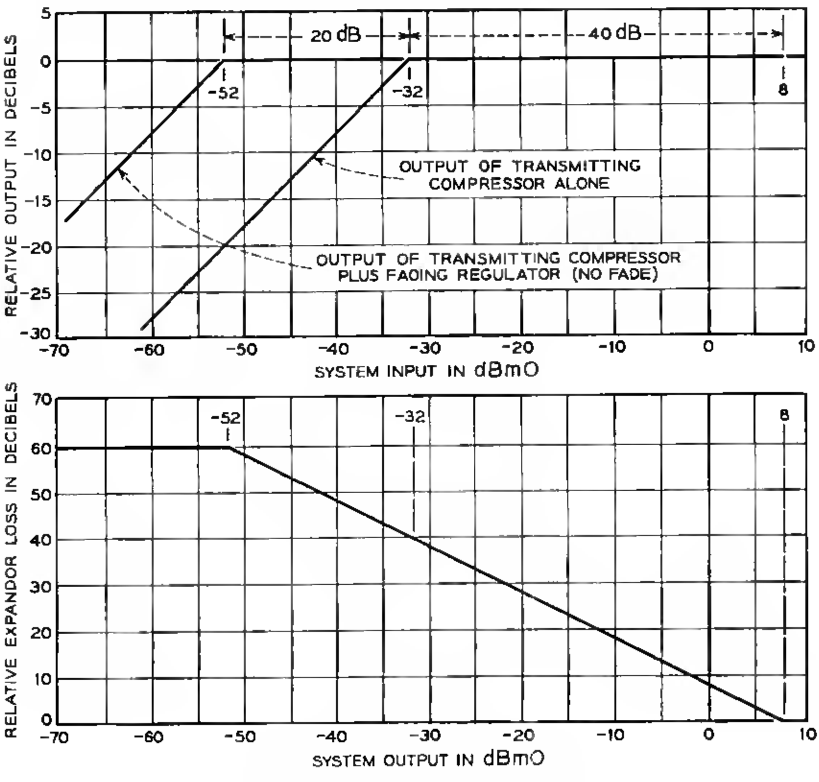

Fig. 13 — Static operating ranges.

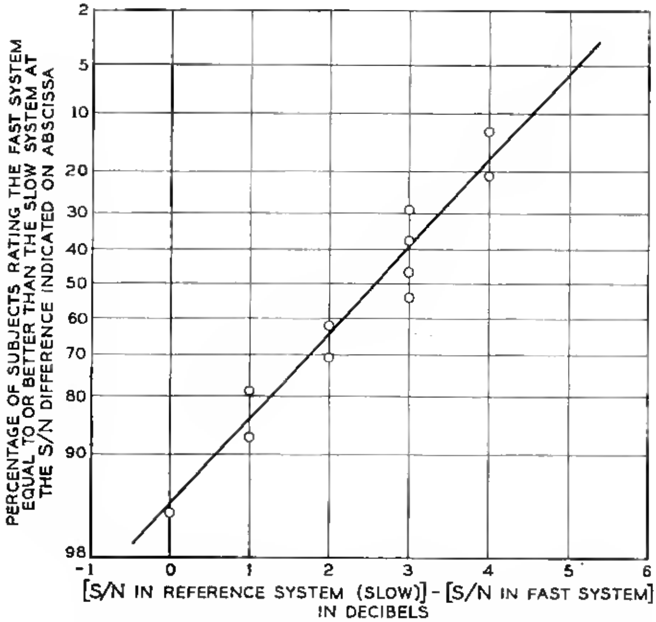

Fig. 14 — Distribution of S/N differences for equal subject preference.

parison of a fast and slow CNL compandor (RC = 5 msec; RC = 75 msec) was made to determine the extent of the change in noise quieting solely related to speed. To this end, the FM control channel was made perfect (replaced by a dc connection), noise was added to the speech path, and recordings were made. The slow system was taken as the reference system, and an approximate 10-dB compressed speech-to-noise ratio maintained. The fast system was recorded with a variable speech-to-noise ratio over a range of 6 to 12 dB. Direct comparisons were made between the two sets of recordings by 12 test subjects. Each subject determined the value of variable noise (in the fast system) which in his judgment made the fast and slow systems equivalent in quality. The reference value of 10 dB was chosen because it is the approximate point at which the CNL system quieting would be most effective on a circuit that was noisy but still usable. Fig. 14 shows the resulting distribution of S/N differences between fast and slow CNL compandors where the subjects indicated equal preference. For the median subject, the fast system gave a 2½-dB improvement. Since the speed-related difference in quieting is strongly affected by the amount of noise present, the 2½-dB advantage will decrease when the S/N improves from the 10-dB test value.

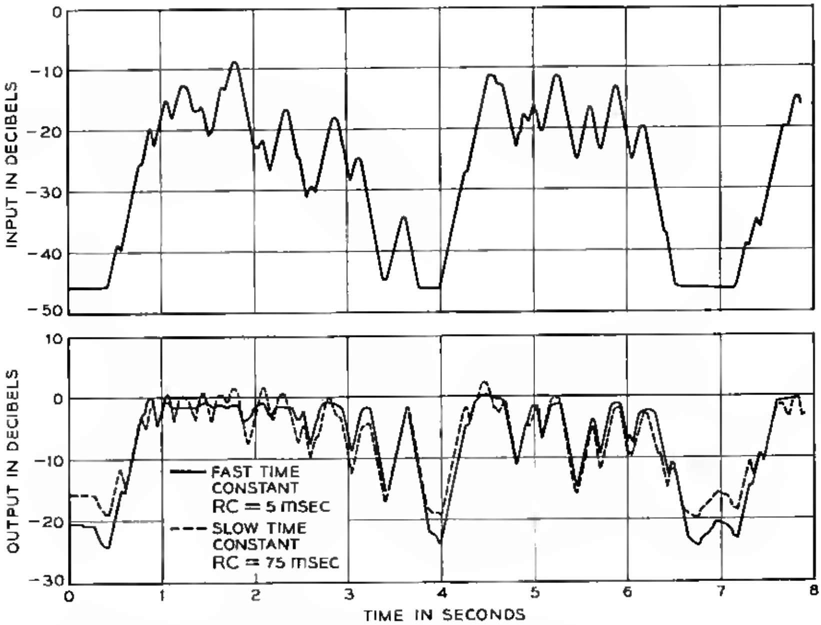

Fig. 15 — Chart record of compressor input and output for fast and slow time constants; recorder integrating time—approximately 100 milliseconds.

The conclusions drawn from the tests were that changes in compandor speed in the range of RC = 5 to 75 msec do not produce a sufficient change in the overall noise improvement to warrant consideration as a major factor in system "trade offs." Since the total noise improvement for the average talker is 20 to 30 dB, the change in performance with speed is of minor consequence.

3.2.6 *Transmitter Loading*

The complete compressor serves to reduce the variations in the speech signal into the radio transmitter. However, residual variations do remain and these variations become greater as the speed of the compandor is reduced. The regulated speech output from the transmitting compressor was examined to determine the extent of the regulation improvement as a result of increasing the speed. The method used in this examination was based upon recognition of current CCIR recommendations.[13] In these recommendations the *mean power* of the speech signal with smoothly read text is used as a criterion of ISB trans-

mitter loading. The *mean power* is defined: "The power supplied to the antenna transmission line by a transmitter during normal operation, averaged over a time sufficiently long compared with the period of the lowest frequency encountered in the modulation. A time of 0.1 sec during which the mean power is greatest will normally be selected." For purposes of comparative measurements, the *mean power* was interpreted as that value obtained by a running measurement of the speech signal power with about a 0.1-sec integrating time and by selection of the value of signal power when the running measurement is greatest, i.e., at the crest of the speech envelope.

CNL output signals were examined using a power measuring chart recorder with about 100-msec averaging time to determine how the *mean power* and the distribution of signal peaks changed as the speed of the transmitting compressor was adjusted to a very fast time constant (RC = 5 msec) and then to a very slow time constant (RC = 75 msec). The input speech and the regulated speech output from the compressor were recorded on a strip chart (see Fig. 15) for each time constant extreme. The strip chart shows the power on a dB scale,

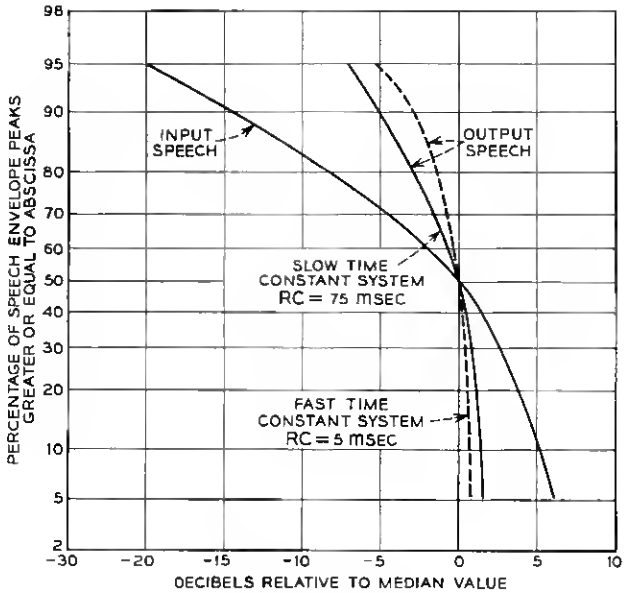

Fig. 16 — Input and output speech envelope peak distribution.

where an output of 0 dB corresponds to the steady state output of the compressor with an input 0 dBm0 1000-Hz tone.

When speech is present, the regulated output signals with both fast and slow time constants are no more than 5 dB apart. More importantly, the slow compressor (with more overshoots expected due to more sluggish regulation) has a mean power that is at most 2 dB higher than the fast compressor. Following accepted practice, the level into the transmitter for RC = 75 msec would then have to be dropped only 2 dB compared to the level for RC = 5 msec. If the peak values for each output are plotted to give a probability distribution (see Fig. 16), the slight difference between the outputs as a function of compressor speed is evident. In Fig. 16 the spread of the speech envelope peak distribution of the input speech is 26 dB. This spread of 26 dB is reduced to 9 dB at the output of the slow compressor and to 6 dB at the output of the fast compressor.

Based on the above results, it was concluded that any otherwise

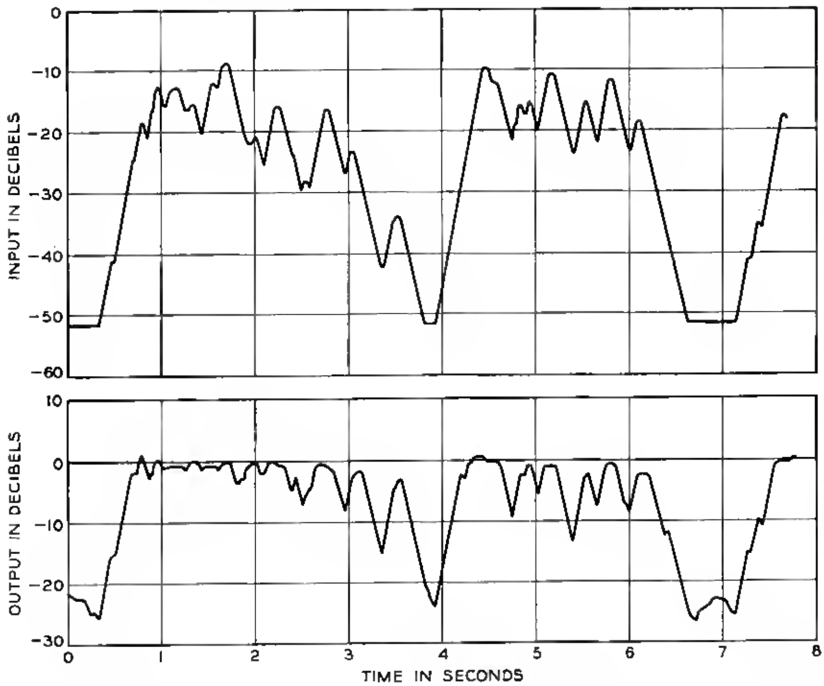

Fig. 17 — Chart record of compressor input and output—RC = 20 milliseconds; recorder integrating time—approximately 100 milliseconds.

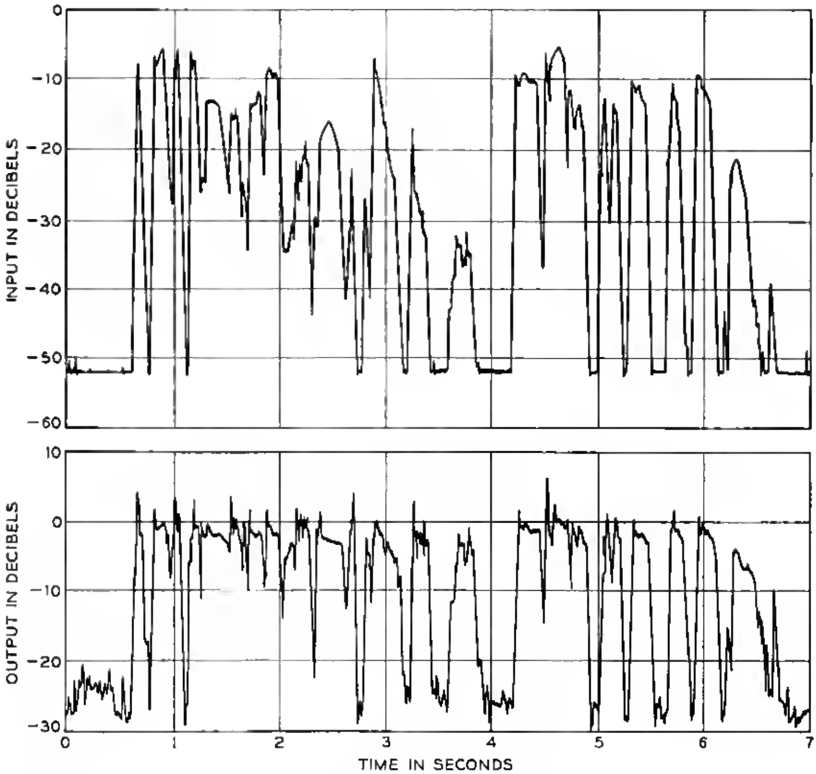

Fig. 18 — Chart record of compressor input and output—RC = 20 milliseconds; recorder integrating time—approximately 10 milliseconds.

proper RC time constant in the range 5 to 75 msec would produce nearly equivalent transmitter loading. The value of $RC_{effective} = 20$ msec used in the trial equipment falls within the above limits and, in fact, produces a load that has almost as little variation as the fast RC = 5 msec compandor. Fig. 17 shows an example of the input and output waveforms for the $RC_{effective} =$ 20-msec system.

To verify that the accepted 100-msec averaging time used in this investigation has not obscured any very short term effects, an additional chart recording was made of system input and output with a 10-msec integrating time. This record, a portion of which is reproduced in Fig. 18, shows that while some overshoots remain, their magnitude and duration are not expected to produce unwanted effects in the radio transmitter.

IV. ACKNOWLEDGMENT

The authors are grateful for the contributions of many persons toward successful implementation and trial of the experimental CNL system, particularly those of Mr. A. E. Donkin.

APPENDIX A

Gain Relations in Single- and Two-Stage Compandors

Application of conventional compandors[14] to conventional wire, cable, and radio relay circuits produces level characteristics such as those shown in Fig. 19. At the compressor, low-level inputs are amplified

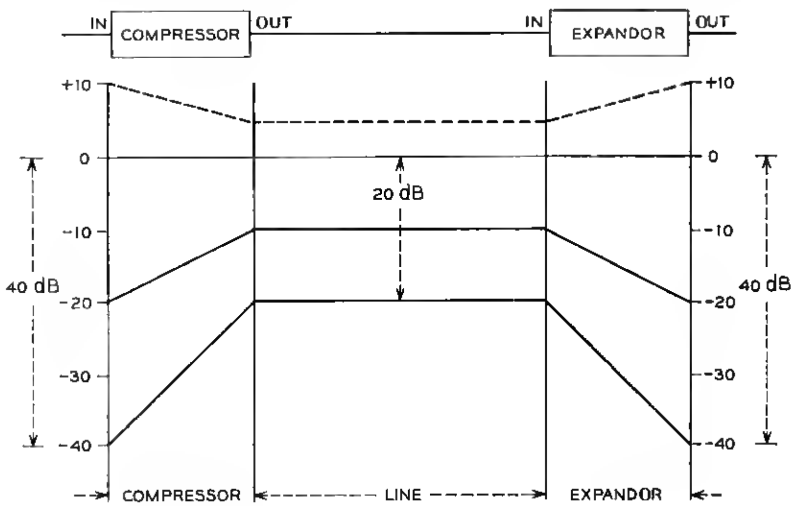

Fig. 19 — Example of 2:1 compandor action.

more than high-level inputs; correspondingly, at the expandor, low-level outputs result from more attenuation than that given to high-level outputs. (Note: certain very high inputs are attenuated to reduce overload.) As shown in the example of Fig. 19, conventional syllabic compandors halve the volume range between compressor and expandor; thus, a 0 to −40-dB range becomes a 0 to −20-dB range in the transmission portion. These compandors are known as 2:1 compandors;* the 2:1 characteristic is easily obtained as will be shown below.

The compressor portion of the compandor consists of only four simple elements (Fig. 20). The variolosser is a circuit configuration where

* In CCITT documents, this is stated as a compression ratio of 2.

the attenuation ratio in the speech path is *inversely* proportional to the unidirectional control current. The variolosser equation is

$$e_1 = k_1 \frac{e_{in}}{i_c}, \tag{8}$$

where e_{in} is the amplitude of the input envelope. The control rectifier acts as an envelope detector, producing control signals with a spectrum extending from 0 Hz to approximately 100 Hz.* The feedback path directs the control signal to control the variolosser attenuation. The 2:1 compression effect results because the control signal developed

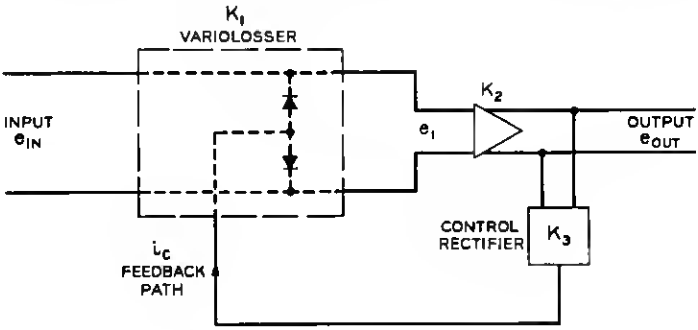

Fig. 20 — Elements of 2:1 compressor.

from the output is fed back to control a gain adjusting variolosser, which in turn changes the output. The fixed gain amplifier is added to the speech path for level adjusting purposes. The amplitude of the output envelope is

$$e_{out} = k_2 e_1 . \tag{9}$$

The control current is related to the output envelope by

$$i_c = k_3 e_{out} \tag{10}$$

from which

$$e_{out} = \frac{k_1 k_2 e_{in}}{i_c} \tag{11}$$

$$e_{out} = \frac{k_1 k_2 e_{in}}{k_3 e_{out}} \tag{12}$$

$$e_{out}^2 = \frac{k_1 k_2}{k_3} e_{in} . \tag{13}$$

* The upper cutoff frequency is not a clearly defined point, depending on the characteristics of the speech signal and the type of rectifier smoothing.

Taking logarithms

$$20 \log e_{\text{out}} = 10 \log e_{\text{in}} + 10 \log \frac{k_1 k_2}{k_3} \tag{14}$$

$$10 \log e_{\text{out}} = \tfrac{1}{2}(10 \log e_{\text{in}}) + K \tag{15}$$

$$E_{\text{out}}(\text{dB}) = \tfrac{1}{2}E_{\text{in}}(\text{dB}) + K'. \tag{16}$$

The output volume changes 1 dB for every 2-dB change in input volume. The foregoing analysis is based upon steady state conditions. The transient response (or time constant) of the compressor is determined by an RC network in the control rectifier. The transient response of compandors and the CNL system is discussed in the analysis section of the body of the paper.

The expandor portion of the compandor consists of five simple elements (Fig. 21). In this case, the variolosser is a circuit configuration where the attenuation ratio in the speech path is *directly* proportional to the unidirectional control current. The control rectifier acts as an envelope detector in the same way as in the compressor, but in this case, the control signal is applied in a forward acting manner rather than in a feedback manner. The variolosser equations

$$e_1 = k_6 i_e e_{\text{in}} \tag{17}$$

$$e_{\text{out}} = k_5 e_1 = k_5 k_6 i_e e_{\text{in}} \tag{18}$$

are then manipulated as in the case of the compressor and the following result is obtained:

$$e_{\text{out}} = k_3 k_4 k_5 k_6 e_{\text{in}}^2 \tag{19}$$

$$E_{\text{out}}(\text{dB}) = 2E_{\text{in}}(\text{dB}) + K''. \tag{20}$$

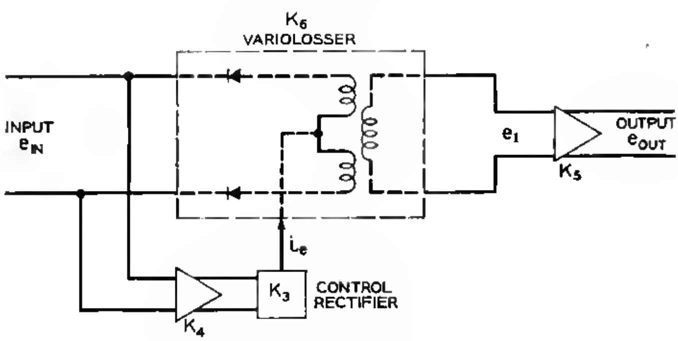

Fig. 21 — Elements of 2:1 expandor.

The output of the expandor changes 2 dB for every dB change in the input, thus undoing the effect of the compressor.

Idealized input-output characteristics of the conventional compandor are shown in Fig. 22. When installed at each end of a circuit, the output will be a true replica of the input if the circuit is distortionless. Since the expandor control signal is derived from the output of the transmission path at the input to the expandor, examination of the

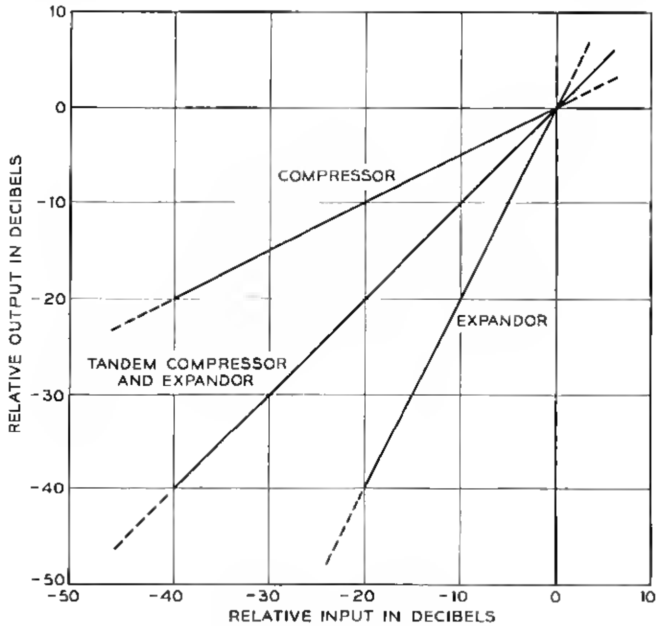

Fig. 22 — Input/output characteristic of 2:1 compandor.

characteristics of Fig. 22 verifies that gain variations (with time) of the medium will be magnified in dB by a factor of two. For this reason, conventional compandors have not found application in HF radio service.*

The CNL compandor arrangement (Fig. 23) solves the problem of net loss variations while providing essentially constant output from the transmitting compressor. Constant output is obtained by adding a second slave variolosser controlled by a replica of the regular con-

* Compandors were used for the first transatlantic radiotelephone circuit[15] (long wave) where the net loss was relatively stable.

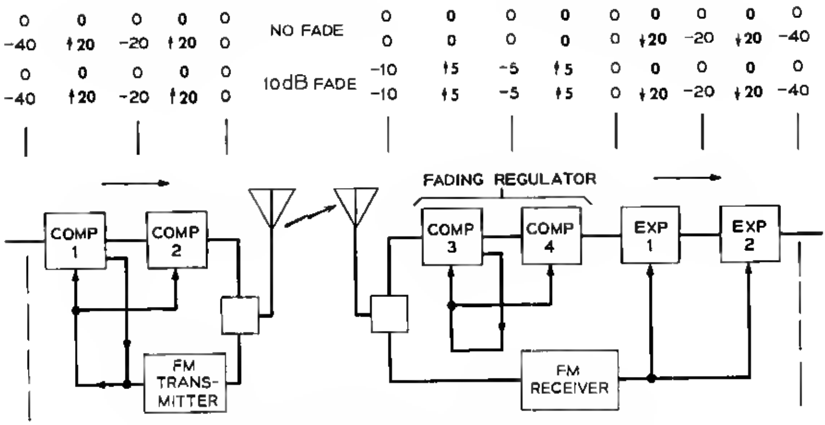

Fig. 23 — Example of CNL system gains and losses.

trol signal present in the feedback path of the normal first stage compressor. Examination of Fig. 22 shows that for any input in the range of compression, a conventional compressor provides a variolosser control signal so as to insert an amount of gain equal to one-half the difference (in dB) between the "unaffected level" (or input level at which no gain or loss occurs in the compressor or expandor) and the input signal level. The slave variolosser gain is also controlled to be one-half the difference between the unaffected level and the input signal level, so the sum of the gains of two stages is sufficient to raise the level of the input signal to the unaffected level for any value of input signal in the range of compression. An example showing the CNL two-stage compressor gains and levels for two inputs is given in Fig. 23.

At the receiving terminal, expandors are used to restore the speech distribution. Two normal expandor variolossers are used, since each can accomplish the inverse of the action of one of the compressor variolossers. However, the local control arrangements at the expandors can not obtain control information from the input signal (which is nominally constant for all speech *in the absence of fading*). Reexamining the compressor and expandor equations

$$\text{compressor gain} = \frac{e_{\text{out}}}{e_{\text{in}}} = \frac{k_1 k_2}{i_c} \tag{21}$$

$$\text{expandor gain} = \frac{e_{\text{out}}}{e_{\text{in}}} = k_5 k_6 i_e \ . \tag{22}$$

Thus, if a compressor and expandor are in tandem, their net gain is constant if

$$i_c \equiv i_e \ . \tag{23}$$

A suitable separate control channel, which supplies each expandor with an i_e that is a replica of i_c, assures this identity.

To maintain a constant net loss when there is fading in the transmission path, the CNL system uses a fading regulator that is similar to the transmitting compressor, except that the fading regulator has no corresponding expandor. The fading regulator action is based on the recognition of the requirement that the output of the radio receiver be constant since the input to the radio transmitter is controlled to a constant value by the transmitting compressor. If the output of the radio receiver is less than the nominally constant value, it is assumed that a fade has occurred.

The fading regulator inserts sufficient gain to raise the output to the nominal nonfade value required by the expandor. Fig. 23 shows the gains and losses with a test tone input for an example without fading and for an example with an assumed flat fade in the medium. In this simple model the overall effect is to provide a nominally constant overall net loss even with a time varying transmission medium.

APPENDIX B

Net Loss Fluctuation Caused by Noise in Control Channel

The effect of noise in the control channel can be obtained by use of the model shown in Fig. 24 (not to scale). A flat band of random noise extends across the speech and subcarrier bands. The bandwidth of the noise affecting the speech directly, $f_b - f_a$, is 2500 Hz. The bandwidth

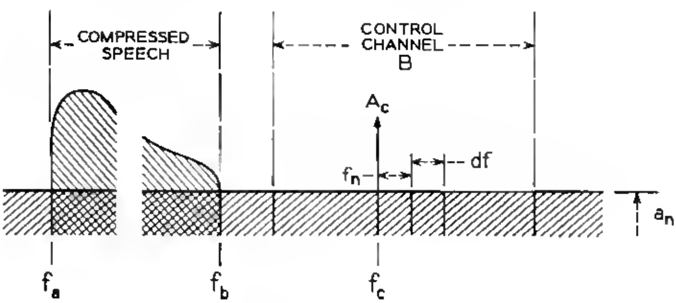

Fig. 24 — Signal and noise spectra.

B is the line, or predetection bandwidth, of the control channel, which is assumed to be defined by an ideal rectangular filter. It is assumed that the peak noise voltage in this band is always less than the peak subcarrier voltage A_c .

In the following analysis, an expression is derived that relates the magnitude of fluctuations in the speech channel loss, to the noise level. The noise level enters this expression implicity in terms of S/N, the speech-to-noise voltage ratio in a 2500-Hz band. It is thus possible to relate judgments concerning S/N and loss fluctuations, respectively.

The flat band of noise of Fig. 24 can be represented by an infinite number of equal-amplitude sinusoids having incommensurable but approximately uniformly spaced frequencies and incoherent phases. The noise in an incremental bandwidth df is represented by a single sinusoid whose mean-square voltage is equal to that of the noise. The mean-square noise voltage in df is just $a_n^2 df$, where a_n is the noise voltage density constant in rms volts per unit square root bandwidth. Thus, if A_n denotes the peak amplitude of the noise component,

$$\frac{A_n^2}{2} = a_n^2\, df. \tag{24}$$

The superposition of the subcarrier and a noise component at a frequency $f_n = \omega_n/2\pi$ relative to the subcarrier can be represented by the phasor diagram (Fig. 25). The peaks of the subcarrier and noise

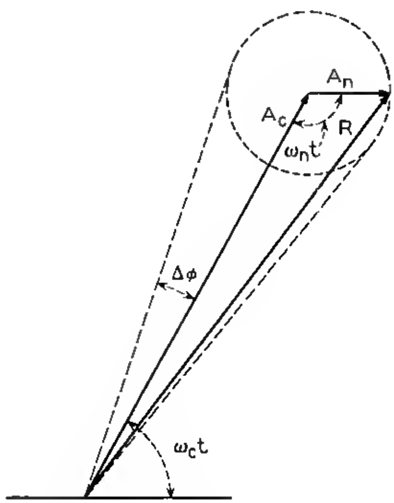

Fig. 25 — Phasor diagram, carrier and noise.

components are denoted by A_c and A_n, respectively. It is assumed that for $A_n \ll A_c$ the resultant angle modulation of the subcarrier produced by the noise will be equal to the summation of the modulations that would be produced by each noise component separately. Except for noise-produced modulation, it will be assumed that the subcarrier is otherwise unmodulated and centered in the band. Amplitude variations may be ignored on the basis that the subcarrier frequency demodulator is ideal.

Rotation of the phasor A_n about A_c causes the resultant R to oscillate about A_c with a peak phase deviation

$$\Delta\varphi = \tan^{-1}\frac{A_n}{A_c} \cong \frac{A_n}{A_c}. \tag{25}$$

It can be shown[16] that if $A_n \ll A_c$ as assumed, $\varphi(t)$ is essentially sinusoidal and equivalent to pure phase modulation of the form

$$M(t) = A_c \sin(\omega_c t + \Delta\varphi \sin pt + \psi),$$

where the modulating signal is a sinusoid of arbitrary radian frequency p. By differentiation of arg $M(t)$, the instantaneous frequency is obtained:

$$\begin{aligned} \omega_i &= \frac{d}{dt}(\omega_c t + \Delta\varphi \sin pt + \psi) \\ &= \omega_c + p(\Delta\varphi)\cos pt. \end{aligned} \tag{26}$$

The instantaneous frequency deviation produced by the noise component at the frequency ω_n relative to the subcarrier can be deduced from (26) by using (25) and setting $p = \omega_n$.

$$\text{Inst. freq. dev.} = \omega_n \frac{A_n}{A_c} \cos \omega_n t, \qquad \text{radians per sec.} \tag{27}$$

The mean-square frequency deviation due to a single noise component is thus

$$\begin{aligned} \langle(\text{freq. dev.})^2\rangle_{av} &= \frac{1}{2}\left(\omega_n \frac{A_n}{A_c}\right)^2, \qquad (\text{radians/sec})^2 \\ &= \left(\frac{A_n^2}{2}\right)\left(\frac{f_n}{A_c}\right)^2, \qquad \text{Hz}^2. \end{aligned} \tag{28}$$

To obtain the incremental mean-square frequency deviation $\langle d\delta^2\rangle_{av}$ due to noise in the band df at a frequency f_n relative to the carrier, (24) is substituted into (28):

$$\langle d\delta^2\rangle_{av} = \left(\frac{a_n f_n}{A_c}\right)^2 df, \qquad \text{Hz}^2. \tag{29}$$

If a discriminator sensitivity of 1 volt per Hz is assumed for convenience, the integral of $\langle d\delta^2 \rangle_{av}$ over the line bandwidth may be equated to the positive-frequency discriminator output mean-square noise voltage spectrum integral:

$$\int_{f_c - B/2}^{f_c + B/2} \langle d\delta^2 \rangle_{av} = \int_0^{B/2} G_n(f)\, df, \tag{30}$$

where

$$G_n(f) = \begin{cases} 2\left(\dfrac{a_n f}{A_c}\right)^2, & \mathrm{V}^2/\mathrm{Hz}, \qquad 0 \leqq f \leqq B/2 \\ 0, & \qquad\qquad f > B/2. \end{cases} \tag{31}$$

The noise spectrum at the output of the single-pole post detection filter is

$$F_n(f) = |\, Y(f)\,|^2\, G_n(f), \qquad \mathrm{V}^2/\mathrm{Hz}, \tag{32}$$

where

$$|\, Y(f)\,| = [1 + (f/f_2)^2]^{-\frac{1}{2}} \tag{33}$$

is the voltage transfer function of the filter and $f_2 = 1/2\pi R_2 C_2$ is the frequency at which the response is down 3 dB. Substituting (31) and (33) into (32) and integrating over half the line bandwidth gives the total mean-square noise voltage output of the control channel that acts on the expandor variolosser:

$$\int_0^{B/2} F_n(f)\, df = 2\left(\frac{a_n}{A_c}\right)^2 \int_0^{B/2} \left[\frac{f^2}{1 + (f/f_2)^2}\right] df, \qquad \mathrm{V}^2. \tag{34}$$

The effect of the ANTILOG circuit (see Fig. 7) on the noise output of the control channel involves a nonlinear transformation of a random process that is difficult to handle mathematically, hence the ANTILOG circuit has thus far been ignored. In spite of this omission, the approximate results derived herein were in good agreement with measurements.

It is to be noted, however, that because of the ANTILOG circuit the translation from a frequency deviation at the FM receiver input to a change in expandor variolosser loss in dB is linear with a slope (or deviation/loss ratio) of 2 Hz per dB. By referring the control channel noise output voltage, (34), back to the discriminator input, an equivalent frequency deviation can be obtained; this deviation, divided by the deviation/loss ratio, gives the fluctuation in expandor loss directly in dB. Accordingly, when this mean square deviation, denoted by $\langle \delta^2 \rangle_{av}$, is equated to the right side of (34) (on the basis of the assumed dis-

criminator sensitivity of 1 volt per Hz) and the integral is evaluated, we have:

$$\langle \delta^2 \rangle_{av} = 2\left(\frac{a_n}{A_c}\right)^2 (f_2)^3 \left[\frac{B}{2f_2} - \tan^{-1}\left(\frac{B}{2f_2}\right)\right], \qquad \text{Hz}^2. \tag{35}$$

In the CNL system, B is 180 Hz. Substituting this value into (35) and taking the square root to get rms, the result is

$$\delta_{rms} = \frac{a_n}{A_c/\sqrt{2}} (f_2)^{\frac{3}{2}} \left[\frac{90}{f_2} - \tan^{-1}\frac{90}{f_2}\right]^{\frac{1}{2}}, \qquad \text{Hz}. \tag{36}$$

This result is next recast in terms of S/N, the ratio of the rms compressed speech voltage to the rms noise voltage in the 2500-Hz speech band. First, note that

$$\frac{A_c}{\sqrt{2}} = \text{rms subcarrier voltage} = \frac{S}{2}, \tag{37}$$

since the subcarrier is transmitted 6 dB below the speech in the CNL system; note also that the rms noise voltage

$$\begin{aligned} N &= \left[\int_0^{2500} a_n^2\, df\right]^{\frac{1}{2}} \\ &= 50\, a_n\,. \end{aligned} \tag{38}$$

Then substitution of (37) and (38) into (36) yields

$$\delta_{rms} = \frac{(f_2)^{\frac{3}{2}}}{25\ \text{S/N}} \left[\frac{90}{f_2} - \tan^{-1}\left(\frac{90}{f_2}\right)\right]^{\frac{1}{2}}, \qquad \text{Hz}. \tag{39}$$

Dividing (39) by the deviation/loss ratio, 2 Hz/dB gives the expandor loss error in dB rms. The result can be expressed in dB peak-to-peak (by ignoring peaks in excess of 3 times rms) if (39) is multiplied by 6. When (39) is modified by these two factors, the result is

$$\begin{aligned} \text{Loss error} &= (\tfrac{1}{2})(6)\ \delta_{rms} \\ &= \frac{3(f_2)^{\frac{3}{2}}}{25\ \text{S/N}} \left[\frac{90}{f_2} - \tan^{-1}\left(\frac{90}{f_2}\right)\right]^{\frac{1}{2}}, \text{ dB peak-to-peak.} \end{aligned} \tag{40}$$

Fig. 8 is a plot of the loss error versus 20 log (S/N) with f_2 as a parameter.

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